METHOD FOR NEAR OPTIMAL JOINT CHANNEL ESTIMATION AND DATA DETECTION FOR COFDM SYSTEMS

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5   This application claims the benefit of priority of Provisional Application No. 60/241,195, filed on October 19, 2000, and of Provisional Application No. 60/180,799, filed on February 7, 2000, and is related by subject matter to U.S. Patent Application No. XX/XXX,XXX filed February 1, 2001, entitled "System for Near Optimal Joint Channel Estimation and Data Detection for COFDM Systems," by the inventors of the present

10   application filed concurrently herewith.

Field of Invention

The invention relates generally to communications and particularly to a method and apparatus for near optimal joint channel estimation and data detection to improve channel tracking and, thus, improve link robustness.

15   ## Background of the Invention

The rapid growth in the use of the Internet and the increasing interest in portable computing devices have triggered the desire for high-speed wireless data services. One of the more promising candidates for achieving high data rate transmission in a mobile environment is Orthogonal Frequency Division Multiplexing (OFDM), which divides the wide signal

20   bandwidth into many narrow-band subchannels, which are transmitted in parallel. Each subchannel is typically chosen narrow enough to eliminate the effects of delay spread. Coded OFDM (COFDM) systems, which combine both OFDM and channel coding techniques, are able to improve the performance further by taking advantage of frequency diversity of the channel.

Though both differential and coherent demodulation can be applied in a COFDM system, the latter leads to a performance gain of 3 to 4-dB in signal-to-noise ratio (SNR) with accurate channel estimation. Channel estimation techniques realized by a frequency-domain filter using Fast Fourier Transform (FFT), followed by time-domain filters for a COFDM 5 system with Reed-Solomon (RS) coding have been proposed. These channel estimation techniques, while good, did not provide the near optimal channel estimation required for data-decoding with improved channel tracking capability for reliable link performance even under high user mobility and/or high RF carrier frequency.

Summary of the Invention

10 The physical layer configuration is shown in Fig. 1. At the transmitter, the encoded data stream is sent to an OFDM transmission branch. The data stream may, for example, be convolutionally encoded. The encoded data stream may then be optionally interleaved. If the encoded data stream is interleaved in the transmitter then the receiver must correspondingly deinterleave the data stream. After interleaving, the transmitter modulates the encoded data 15 stream. By way of example, QPSK modulation is used. The signal is then subjected to inverse Fast Fourier Transformation and transmitted, in the present invention over the air.

Correspondingly, a receiver accepts multicarrier transmitted signals and subjects these received signals to Fast Fourier Transformation. The transformed signals are concurrently fed into a channel estimator and demodulators. The demodulated signals are 20 combined in a maximum ratio combiner, optionally deinterleaved and decoded.

An OFDM signal is divided into a number of subchannels. By way of example, an OFDM signal bandwidth is divided into 120 6.25-kHz subchannels with QPSK modulation on each subchannel. At the receiver, the demodulated signals from two receiving branches

are combined using maximal ratio combining and then decoded. With a symbol period of 200 μs (including a 40-μs guard interval) and 1/2-rate coding, a maximum information rate of 600 kbps can be achieved in a 750-kHz bandwidth (about 800 kHz including guard bands). The information rate is calculated by dividing the 120 subchannels (tones) by the 200 μs period to obtain 600 kbps.

For purposes of example for the present invention, 1/2-rate convolutional codes (CC) are considered. The results with 1/2-rate Reed-Solomon (RS) code based on Galois-Field (64) (GF(64)) are compared. The size of a code word is the same as that of an OFDM block (an OFDM symbol of 200 μs and 120 subchannels). To achieve coding gain with inherent frequency diversity in OFDM, a simple interleaving scheme is applied. For both RS and CC cases, the first 120 bits of a code word are assigned to the in-phase component and the rest to the quadrature component. To gain additional randomness within a code word for the CC case, each 120-bit group is interleaved over subchannels by an 11-by-11 block interleaver (without the last bit).

In the simulations, the wireless channel, as a Rayleigh-fading channel, with a two-ray multipath delay profile is modeled. Good performance for impulse separation as high as 40-μs can be achieved; a 5-μs impulse separation in the numerical results is considered.

For the performance with respect to channel variations, maximum Doppler frequency up to 200 Hz, which is reasonable for most vehicular speeds, for a possible RF carrier frequency around 2 GHz is considered. To demonstrate the advantage of the proposed joint detection methods, results at a maximum Doppler frequency as high as 500 Hz corresponding

to a scenario in which the wireless system uses a higher carrier frequency, e.g. 5 GHz are presented.

In the medium access control (MAC) layer, a frequency reuse is considered with dynamic resource management, e.g., Dynamic Packet Assignment (DPA), to achieve high spectral efficiency for packet data access.

A simple analysis to highlight the ideal or optimal joint channel estimation and maximum likelihood (ML) decoding scheme indicated in Fig. 1A for the case of $M = 2$ receiving antennas is now presented.

At a diversity receiver, the signal from the mth antenna at the kth subchannel and the nth block can be expressed as

$$x_{m,n,k} = h_{m,n,k} a_{n,k} + w_{m,n,k}, \qquad (1)$$

where $a_{n,k}$, $h_{m,n,k}$ and $w_{m,n,k}$ are the transmit signal, channel response and additive Gaussian noise, respectively.

For convolutional codes, because the size of a code word is the same as that of the OFDM block, (1) can be rewritten as

$$\mathbf{x}_{m,n} = \mathbf{H}_{m,n}\mathbf{c}_n + \mathbf{w}_{m,n}, \qquad (2)$$

where, if there are K_f subchannels,

$$\mathbf{H}_{m,n} = diag(h_{m,n,1}, h_{m,n,2}, ..., h_{m,n,K_f}),$$

$\mathbf{c}_n$ is the transmitted code word at time epoch n, and the rest of the vectors are similarly defined.

Assume that the number of code words is N, we introduce the following notations,

$$\mathbf{c} = [\mathbf{c}_1^T, \mathbf{c}_2^T, ..., \mathbf{c}_N^T]^T,$$

$$\mathbf{H}_m = diag(\mathbf{H}_{m,1}, \mathbf{H}_{m,2}, ..., \mathbf{H}_{m,N}),\qquad(3)$$

$$\mathbf{x}_m = [\mathbf{x}_{m,1}^T, \mathbf{x}_{m,2}^T, ..., \mathbf{x}_{m,N}^T]^T.$$

At the receiver, the objective is to solve a maximum likelihood (ML) problem

$$\hat{\mathbf{c}} = \arg\min_{\mathbf{c}}[\min_{\mathbf{H}_m}\sum_m \|\mathbf{x}_m - \mathbf{H}_m\mathbf{c}\|^2],\qquad(4)$$

with a constraint on channel response

$$L(\mathbf{H}_m) = 0,\qquad(5)$$

5   where $L()$ is a constraint function. In a wireless environment, this constraint can be simplified to be

$$\sum_{l=-K_m}^{K_m} \mathbf{B}_{n,l}\mathbf{d}(\mathbf{H}_{m,n-l}) = 0,\qquad(6)$$

where the length of the channel memory is K_m OFDM symbol durations, $\mathbf{B}_{n,l}$ are coefficients determined by the correlation between channel responses at the time epochs n and n-l, which is a function of the Doppler spectrum of the channel, and $\mathbf{d}()$ is a vector

10   function defined by

$$\mathbf{d}(\mathbf{H}_{m,n}) = [h_{m,n,1}, h_{m,n,2}, ..., h_{m,n,K_f}]^T.$$

The optimal solution of this ML problem can be obtained by exhaustive search. It requires solving the mean square error (MSE)

$$MSE(\mathbf{c}) = \min_{\mathbf{H}_m} \sum_m \| \mathbf{x}_m - \mathbf{H}_m \mathbf{c} \|^2, \qquad (7)$$

for any possible $\mathbf{c}$ with the channel constraint (6). Then,

$$\hat{\mathbf{c}} = \arg\min_{\mathbf{c}} MSE(\mathbf{c}). \qquad (8)$$

5   After obtaining $MSE(\mathbf{c})$, the corresponding channel estimate $\mathbf{H}_m(\mathbf{c})$ can be found. Consequently, the optimal approach for estimating channel response requires the knowledge of the entire set of $\mathbf{x}$ and $\mathbf{c}$.

Another observation from this ML receiver is that the channel estimation results $\mathbf{H}_m$ is not a direct output of the detection process and hence, channel estimation which calculates

10    $\mathbf{H}_m$ explicitly may not be necessary in theory. However, for other required parameter estimation, such as timing and frequency synchronization, a known data sequence is usually transmitted in the beginning of a group of OFDM blocks. This known data sequence, also called a synch word or a unique word, can be used as a training sequence in (7) to obtain initial channel estimate explicitly without resorting to blind detection. This initial channel

15   characteristic is helpful for solving this ML problem with better numerical stability and tracking property. This initial channel estimation can be easily solved in the frequency domain by first taking FFT as shown in Fig. 1A.

One related method and system is Serial Number 09/089,862, METHOD AND APPARATUS FOR CHANNEL ESTIMATION FOR MULTICARRIER SYSTEMS, which

was filed June 3, 1998, and is commonly held and incorporated herein by reference. The near optimal joint channel estimation and data detection method and system of the present invention was born from the research that resulted in that application and the subsequent determination that improvements could be made in the channel estimation.

The sub-optimal approach of the related system and method is now outlined. Because of the formidable complexity of the optimal ML receiver, some sub-optimal solutions are widely used in practice. The related sub-optimal solution is to divide the ML problem into two parts, channel estimation and coherent decoding. Then, the problem can be solved by iteratively estimating channel and decoding in the forward direction (in time).

At a time instant n, given a channel estimate $\hat{\mathbf{H}}_{m,n}$, initially obtained by using training sequence (in the frequency domain dividing the transfer function of the received signal by the transfer function of the known data), the maximum likelihood (ML) problem

$$\hat{\mathbf{c}}_n = \arg\min_{\mathbf{c}_n} \sum_m \|\mathbf{x}_{m,n} - \hat{\mathbf{H}}_{m,n}\mathbf{c}_n\|^2 \qquad (9)$$

can be solved. Then, the reference for channel estimation is

$$\tilde{\mathbf{H}}_{m,n} = \arg\min_{\mathbf{H}_{m,n}} \sum_m \|\mathbf{x}_{m,n} - \mathbf{H}_{m,n}\hat{\mathbf{c}}_n\|^2. \qquad (10)$$

Finally, the estimate for the time instant $n+1$ is obtained by solving a linear constrained equation (6). Considering a stationary channel with fixed maximum Doppler frequency, the coefficients $\mathbf{B}_{n,l}$ are independent of n, and can be written as $\mathbf{B}_l$, which are used as the coefficients in the FIR filter to track channel variations. Consequently, a simplification of (6) with only previous references for prediction-type estimation leads to

$$\sum_{l=1}^{M_L} \mathbf{B}_l \mathbf{d}(\widetilde{\mathbf{H}}_{m,n+1-l}) - \mathbf{d}(\hat{\mathbf{H}}_{m,n+1}) = 0, \qquad (11)$$

where M_l is the number of taps of an FIR filter and $\mathbf{B}_l$ are preset coefficients designed to achieve the minimum mean square error (MMSE) of estimation. This MMSE estimator can be realized by a frequency-domain filter using the Fast Fourier Transform (FFT), followed by an M_l-tap time-domain filter,

$$\mathbf{B}_l = b_l \mathbf{F}^{-1} \mathbf{B}_f \mathbf{F}, \qquad (12)$$

where b_l is the time-domain filter coefficient, $\mathbf{F}$ is the FFT matrix, $\mathbf{B}_f$ is a diagonal matrix, and $\mathbf{F}^{-1} \mathbf{B}_f \mathbf{F}$ is the frequency-domain filter.

The MMSE filter coefficients b_l and $\mathbf{B}_f$ were derived assuming $\hat{\mathbf{c}}_n = \mathbf{c}_n$ for a given set of Doppler frequency and delay spread. It is shown that this estimator is robust regardless of frequency or time mismatches. With a low Doppler frequency, it has been shown, that a 5-tap (M_l=5) estimator can successfully predict the channel.

To obtain accurate initial channel estimation, a training OFDM block is sent at the beginning of a transmission, in which $\mathbf{c}_1$ is known to obtain $\widetilde{\mathbf{H}}_{m,1}$. The channel parameters for the new time epochs and the unknown code words can be successively solved in the forward direction (time advance).

The assumption $\hat{\mathbf{c}}_n = \mathbf{c}_n$ cannot be always guaranteed; an incorrectly detected code word introduces wrong channel estimation and hence, can cause a wrong detection of the successive code words. This error propagation is the dominant impairment of the link performance at high SNR. In order to alleviate the problem of error propagation, a training

OFDM block is inserted every N_t block. In the simulations presented herein, $N_t = 10$ is considered.

In Fig. 2, the performance of the coherent reception with RS code and convolutional codes for 40-Hz maximum Doppler frequency is shown. The convolutional codes are shown with different constraint length (K) ranging from 3 to 9. The performance of convolutional codes is substantially better than that of the RS code. In order to achieve a Word Error Rate (WER) of 10^{-2}, the K=9 CC needs 4 dB lower SNR than the RS code. In the use of WER in the present invention, it is assumed that a word is a codeword. Moreover, the performance of the K=9 CC with channel estimation is very close to the one with the ideal channel information.

In Fig. 3, the performance at 200-Hz Doppler is shown. In comparison with the RS code, the CC's are still superior although the degradation with respect to the idealized case is higher due to poorer channel tracking. In fact, an error floor at the high SNR region exists due to tracking errors.

In Fig. 4, the performance of the K=9 CC with different maximum Doppler frequencies is shown. With a low Doppler frequency, a 5-tap (M_t =5) estimator used here can successfully predict the channel and the performance is very close to that with idealized channel estimation. However, when the fading is relatively fast, it is difficult to estimate the channel correctly and the WER floors on the order of 10^{-3} can be clearly observed for at a maximum Doppler frequency of 200 Hz. That is, it was found that the original method works well in slow fading but degrades significantly in fast fading.

Fig. 1B depicts a block diagram for a baseband receiver. It is simplified to the extent that only channel estimator and decoder portions of the receiver are depicted. The decoder further comprises a plurality of demodulators, as well as a maximum ratio combiner, an optional deinterleaver (to match an optional interleaver in the transmitter) and a Viterbi

5 decoder, all depicted in Fig. 1A. However, Fig. 1B depicts the structure and corresponding connections of the receiver depicted in Fig. 1A. Note, all $\mathbf{x}$, $\hat{\mathbf{c}}$, $\hat{\mathbf{H}}$, $\tilde{\mathbf{H}}$ are complex. Although it may seem strange to have a complex codeword, $\hat{\mathbf{c}}$, it is very natural to treat coding and modulation as a whole like spatial-temporal coding as discussed in "Space-time codes for high data rate wireless communication: performance criterion and code

10 construction", by V. Tarokh, N. Seshadri, and A. Calderbank, published in *IEEE Trans. Info. Theory*, vol. 44, no. 2, pp. 744-765, Mar. 1998 or coded modulation "Channel coding with multilevel/phase signals," by G. Ungerboeck, published in *IEEE Trans. Info. Theory*, vol. IT-28, no. 1, pp. 55-67, Jan. 1982.

At a time instant n, the channel estimator unit has two tasks. One is to produce the

15 channel estimates of current time instant, $\hat{H}_{1,n}, \hat{H}_{2,n}, ...$ with its input $x_{1,n}$, $x_{2,n}$, ... and feedback $\tilde{H}_{1,n-M_l}, \tilde{H}_{2,n-M_l}, ..., \tilde{H}_{l,n-1}, \tilde{H}_{2,n-1,...}$ by equation (11). The other task is to produce references $\tilde{H}_{1,n}, \tilde{H}_{2,n}, ...$ for estimate processing in the next time instant by equation (10) when $\hat{c}_n$ is available. At a time instant n, when $\hat{H}_{1,n}, \hat{H}_{2,n}, ...$ are available, the decoder unit produces decoded information $\hat{c}_n$ by (9).

20 The flowchart of this method is shown in Fig. 1C. The related channel estimation method is first initialized. Transmitted signals are received. A determination is then made as

to whether the received block is a training block. If the received block is a training block then $\hat{c}_n$ is known and

$$\widetilde{\mathbf{H}}_{m,n} = \arg\min_{\mathbf{H}_{m,n}} \sum_m \|\mathbf{x}_{m,n} - \mathbf{H}_{m,n}\hat{\mathbf{c}}_n\|^2$$

is calculated. This is a reference for the channel estimation

$$\sum_{l=1}^{M_L} \mathbf{B}_l \mathbf{d}(\widetilde{\mathbf{H}}_{m,n+1-l}) - \mathbf{d}(\hat{\mathbf{H}}_{m,n+1}) = 0$$

which is calculated next. The block number is incremented and a determination is made if the end of the frame has been reached. If the current block is not a training block then $\hat{\mathbf{c}}_n$ is decoded

$$\hat{\mathbf{c}}_n = \arg\min_{\mathbf{c}_n} \sum_m \|\mathbf{x}_{m,n} - \hat{\mathbf{H}}_{m,n}\mathbf{c}_n\|^2$$

is calculated before calculating the reference and channel estimation.

It should be clear from the foregoing that there is room for improvement between the prior art and optimal joint channel estimation and data detection. An object, therefore, of the present invention is to improve the joint channel estimation. This will have the effect of reducing the impact of noise as well as reducing decoding errors. Thus, overall system performance will be improved.

It is a further object of the present invention to provide a method and system that are robust even in light of a mismatch between Finite Impulse Response (FIR) coefficients and the true channel.

Another object of the present invention is to provide a method and system with improved channel tracking capability, resulting in reliable link performance even under high user mobility and/or high RF carrier frequency. With improved link performance, data rates

significantly higher than currently available (or even than third generation systems in planning) can be offered to subscribers.

All of the above objects can be achieved nearly optimally even in rapid dispersive fading channels.

Brief Description of the Drawings

The present invention will now be described in greater detail with reference to the preferred embodiments of the invention, given only by way of example, and illustrated in the accompanying drawings, in which same elements are numbered the same to the extent possible:

Fig. 1A is the physical configuration of a system for transmitting and receiving signals.

Fig. 1B is a block diagram for a baseband receiver at time instant n.

Fig. 1C is a flowchart for a related sub-optimal channel estimation method.

Fig. 2 is a comparison of RS and convolutional codes, where the maximum Doppler frequency is 40 Hz.

Fig. 3 is a comparison of RS and convolutional codes, where the maximum Doppler frequency is 200 Hz.

Fig. 4 depicts the performance of the K=9 CC at different maximum Doppler frequencies.

Fig. 5 depicts the performance of the iterative approach at 200 Hz maximum Doppler frequencies, where K=9.

Fig. 6A is a flowchart for the iterative processing.

Fig. 6B is a flowchart for the iterative backward processing.

Fig. 7 shows the performance of iterative backward-processing approach at 200 Hz maximum Doppler frequency, where K=9.

Fig. 8 shows the performance of iterative backward-processing approach at different maximum Doppler frequencies, where K=9.

5   Fig. 9 is a comparison of systems having different time-domain FIR estimators, where K=9.

Fig. 10 shows the average retransmission probability as a function of occupancy per sector. Eight slots and three RF carriers (< 2.5 MHz) are reused in every base station, each with three sectors, using Dynamic Packet Assignment (DPA). In the original (non-iterative)

10   method, K=5, 40 Hz Doppler or K=9, 125 Hz Doppler.

Fig. 11 shows the average delay of delivered packets as a function of throughput per base station, each with three sectors. Eight slots and three RF carriers (< 2.5 MHz) are reused in everywhere using DPA. In the original (non-iterative) method, K=5, 40 Hz Doppler or K=9, 125 Hz Doppler.

15   Fig. 12 depicts the average retransmission probability as a function of occupancy per sector. Eight slots and three RF carriers (< 2.5 MHz) are reused in every base station, each with three sectors, using DPA, where K=9, 200 Hz Doppler.

Fig. 13 shows the average delay of delivered packets as a function of throughput per base station, each with three sectors. Eight slots and three RF carriers (< 2.5 MHz) are reused

20   in everywhere using DPA, where K=9, 200 Hz Doppler.

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Detailed Description of the Preferred Embodiments

A method and system using two channel estimation schemes, which improve channel-tracking capability based on simplification of the ideal or optimal detector are now presented.

The present invention uses a COFDM system with convolutional coding (CC). Moreover, the present invention uses two iterative processing techniques using current and future tentative data decisions in the backward direction for channel estimation and the final decoding. This permits joint channel estimation and data decoding with improved channel-tracking capability, resulting in reliable link performance even under high user mobility and/or high RF carrier frequency. System level performance such as retransmission probability and packet delay in a system with aggressive frequency reuse using dynamic packet assignment (DPA) is also presented.

Since the wireless channel is correlated in time, the optimal joint detection requires processing the received signals and the decoded data in the past, current and future epochs. From (10) and (11), it can be seen that the joint detector in the related is sub-optimal because it attempts to predict the channel response at a time instant by only using received signals and decoded information in the past. In order to improve channel estimation by applying the current information, e.g. decoded data and newly received signal, an iterative solution can be applied.

At each time epoch n, after performing (9), (10) and (11), the present invention does not proceed to the time epoch $n+1$ immediately. Instead, the present invention uses the newly generated $\hat{\mathbf{H}}_{m,n+1}$, which is derived based on the current information, to substitute for the current channel estimate $\hat{\mathbf{H}}_{m,n}$ and the computations in (9), (10) and (11) are repeated. In

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other words, the same procedures are performed twice. This additional iteration allows channel estimation to be performed based on the *current* information.

In Fig. 5, the improvement of the system performance is shown. At each time instant n, if the perfect past channel were known, i.e., $[\tilde{\mathbf{H}}_{m,1},,,\tilde{\mathbf{H}}_{m,n-1}] = [\mathbf{H}_{m,1},,,\mathbf{H}_{m,n-1}]$ in (11), the new iterative processing can perform within .3 dB from the case in which channel is known. On the other hand, even if the perfect past channel information were available, the system performance would still be far from the optimal one with an irreducible error floor in the related sub-optimal solution. The SNR requirement of the new iterative approach at a WER of 10^{-2} is 1.2 dB lower than that of the original one.

More than two iterations were attempted but the improvement was limited. Therefore, it was determined that another possible improvement probably should come from utilizing *future* information as described next.

Following the insights of the optimal approach, by taking into account both the *current and* the *future* information another step is performed. To take advantage of the future information in the channel estimation process, the FIR filter shown in (11) can be processed in the time-reversed fashion, i.e.,

$$\sum_{l=1}^{M_L} \mathbf{B}_l \mathbf{d}(\tilde{\mathbf{H}}_{m,n+l-1}) - \mathbf{d}(\hat{\mathbf{H}}_{m,n-1}) = 0. \qquad (13)$$

Hence, the iterative approach is performed backwards. Here, assume that the iterative approach mentioned in the last sub-section has been already carried out from time instant 1 to N. Therefore, $\mathbf{x}_m$ and $\hat{\mathbf{c}}$ are available with which to perform the iterative approach backward at a later time for tentatively detected previous signals; this method is thus referred

to as *iterative backward-processing* approach. In order to process backward, $\widetilde{\mathbf{H}}_{m,n}$ *is stored when the decoding result is correct (assuming an error detection mechanism is available) and $\hat{\mathbf{c}}$ and $\mathbf{x}_m$ when a decoding error occurs at a time instant n.*

In general, there are three choices to perform the backward processing. The first choice is to process backward after all N OFDM-block data are processed. In reality, it is not a good choice because the channel memory length K_m is not infinite and the delay of this approach may not be acceptable if N is large. A feasible approach is to process backward after having M_I (FIR tap length) consecutive correct code words (OFDM blocks). This ensures that the backward processing will start from a reliable channel estimate. The last choice is to process backward starting at any other points, which is determined by memory and delay requirements. Performance degradation is expected. However, if the backward processing is started from a known training OFDM block, lower degradation can be achieved. With the last two choices, it is only necessary to store M_N ($<< N$ in general) OFDM-block data in the memory.

The *iterative processing* and *iterative backward processing* approaches share the similar baseband processing block diagram (Fig. 1A) with the related/original approach. The only difference is that the feedback is $\widetilde{\mathbf{H}}_{1,n-M_I+1}, \widetilde{\mathbf{H}}_{2,n-M_I+1}, \ldots, \widetilde{\mathbf{H}}_{1,n}, \widetilde{\mathbf{H}}_{2,n}\ldots$ and $\widetilde{\mathbf{H}}_{1,n-M_I}, \widetilde{\mathbf{H}}_{2,n-M_I}, \ldots, \widetilde{\mathbf{H}}_{1,n+1}, \widetilde{\mathbf{H}}_{2,n+1}\ldots$, respectively, when performing iterative processing and iterative backward processing. Their corresponding flowcharts are shown in Fig. 6A and Fig. 6B.

The iterative channel estimation method depicted in Fig. 6A is first initialized. Transmitted signals are received. A determination is then made as to whether the received block is a training block. If the received block is a training block then $\hat{c}_n$ is known and

$$\widetilde{\mathbf{H}}_{m,n} = \arg\min_{\mathbf{H}_{m,n}} \sum_m \|\mathbf{x}_{m,n} - \mathbf{H}_{m,n}\hat{\mathbf{c}}_n\|^2$$

5   is calculated. This is a reference for the channel estimation

$$\sum_{l=1}^{M_L} \mathbf{B}_l \mathbf{d}(\widetilde{\mathbf{H}}_{m,n+1-l}) - \mathbf{d}(\hat{\mathbf{H}}_{m,n+1}) = 0$$

which is calculated next. The block number is incremented and a determination is made if the end of the frame has been reached. If the current block is not a training block then $\hat{c}_n$

$$\hat{\mathbf{c}}_n = \arg\min_{\mathbf{c}_n} \sum_m \|\mathbf{x}_{m,n} - \hat{\mathbf{H}}_{m,n}\mathbf{c}_n\|^2$$

10   is calculated. Then the reference

$$\widetilde{\mathbf{H}}_{m,n} = \arg\min_{\mathbf{H}_{m,n}} \sum_m \|\mathbf{x}_{m,n} - \mathbf{H}_{m,n}\hat{\mathbf{c}}_n\|^2$$

is calculated followed by calculating

$$\sum_{l=1}^{M_L} \mathbf{B}_l \mathbf{d}(\widetilde{\mathbf{H}}_{m,n+1-l}) - \mathbf{d}(\hat{\mathbf{H}}_{m,n+1}) = 0$$

the channel estimation. The method then proceeds to the step after determining that the current block is a training block. This effectively repeats the reference and channel estimation. The reference and channel estimation are repeated in order to improve the calculations with the tentative reference and channel estimation calculations.

The iterative backward channel estimation method depicted in Fig. 6B is first initialized. Transmitted signals are received. A determination is then made as to whether the received block is correct. If the received block is correct then $\hat{c}_n$ is known and

$$\widetilde{H}_{m,n} = \underset{H_{m,n}}{\arg\min} \sum_{m} \|x_{m,n} - H_{m,n}\hat{c}_n\|^2$$

5   is calculated. This is a reference for the channel estimation

$$\sum_{l=1}^{M_L} B_l d(\widetilde{H}_{m,n+l-1}) - d(\hat{H}_{m,n-1}) = 0.$$

which is calculated next. The block number is decremented and a determination is made if the beginning of the frame has been reached. If the current block is correct block then $\hat{c}_n$

$$\hat{c}_n = \underset{c_n}{\arg\min} \sum_{m} \|x_{m,n} - \hat{H}_{m,n}c_n\|^2$$

is calculated. Then the reference

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$$\widetilde{H}_{m,n} = \underset{H_{m,n}}{\arg\min} \sum_{m} \|x_{m,n} - H_{m,n}\hat{c}_n\|^2$$

is calculated followed by calculating

$$\sum_{l=1}^{M_L} B_l d(\widetilde{H}_{m,n+l-1}) - d(\hat{H}_{m,n-1}) = 0.$$

the channel estimation. The current block is then redecoded. The method then proceeds to the step after determining that the current block is correct. This effectively repeats the reference and channel estimation. The reference and channel estimation are repeated in order to

15   improve the calculations with the tentative reference and channel estimation calculations.

In Fig. 7, the performance of this approach is shown for 200 Hz maximum Doppler frequency. In the simulation, the maximum M_N is set to be 200 and this corresponds to 40

ms. However, due to the low error probability in the high SNR region, much shorter storage is required. For instance, the maximum M_N required is about 50 at the 5 dB SNR. It is found that nearly optimal performance is achieved with iterative backward processing.

In Fig. 8, the performance of this approach with different maximum Doppler frequencies is shown. The system still performs well in an environment with maximum Doppler frequency as high as 400 Hz.

By applying the *simple* iterative estimation, the impact of noise is reduced by an additional round of filtering with newly available information. However, it may not eliminate the impact of decoding errors. Using iterative *backward-processing*, there is a chance to re-estimate the channel with fewer decoding errors and hence, better performance is achieved.

It should be noted that the FIR estimator coefficients $\mathbf{B}_l$ in the related sub-optimal method were optimized for the scenario without any iterative process. It should be noted, however, *the detector architecture introduced here can be combined with any filter design for better tracking.*

To show the robustness of the iterative approaches, a simple averaging 5-tap FIR ($b_l \equiv .2$) for the time domain filtering is now considered. As shown in Fig. 9, even with these simple FIR coefficients, the method of the present invention still outperforms the original method with the FIR that was optimized for a particular set of maximum Doppler frequency and delay spread. Therefore, the iterative backward-processing approach of the present invention is relatively robust against the mismatch between the FIR coefficients and the true channel.

Consider a MAC layer configuration to characterize the system-level performance under frequency reuse using the improved detection methods. A simulation system of 36 base stations arranged in a hexagonal pattern is used, each having 3 sectors, with a 20-dB front-to-back ratio and idealized antenna pattern. The same channel can be used everywhere, even in different sectors of the same base station, as long as the Symbol to Interference Ratio (SIR) in the DPA admission process exceeds 7 dB. The simulation of the simultaneous use of the same spectrum by users in different cells results in interference between cells. Interference is treated as if it behaves like noise.

A channel is defined to be a combination of time slot and RF carrier, each consisting of 120 subchannels described previously. By way of example, the results for the case with *3 RF carriers and 8 time slots* are now presented. This occupies a total spectrum of less than 2.5 MHz, including guard bands and other overhead. Each time slot consists of 10 OFDM blocks, i.e., 2 msec. One of these OFDM blocks is used for training, as discussed previously, while an additional block can be allocated for guard time between time slots. In addition, assume that a control slot of duration 4 msec is inserted in the beginning of every frame of 8 traffic slots to enable paging, assignment and pilot transmission that are required for the DPA process. With this conservative assumption of overhead in time and frequency domains, *48 kb/s* (8 OFDM blocks or 960 data bits transmitted in 20 msec) can be delivered using *each time slots*. Once paged, a mobile station (MS) measures the pilot signals to determine the desired traffic slots and reports the list back to the base station (BS). The BS then assigns traffic channel(s) and informs the MS this assignment for traffic packet delivery.

Based on the downlink frame structure, four adjacent BS's form a reuse group and they take turns performing the DPA procedure once every 4 frames. The time-reuse groups in the entire service area are pre-planned in a fixed and repeated pattern.

For the propagation model, the average received power decreases with distance d as d^{-4} and the large-scale shadow-fading distribution is log-normal with a standard deviation of 10 dB. A data-service traffic model, based on wide-area network traffic statistics, which exhibit a "self-similar" property when aggregating multiple sources, was used to generate packets.

Automatic Retransmission reQuest (ARQ) is employed for retransmission when a packet is received in error. A packet in this case is theoretically 8 code words in each time slot, but the error probability is represented by using the WER curves. Since the error probability of the 8 code words in a time slot is highly correlated and additional coding is usually included for the entire packet, this approximation provides reasonable performance estimation for the MAC layer. If a packet cannot be successfully delivered in 3 seconds, which may be a result of traffic overload or excessive interference, it is dropped from the queue. The control messages are assumed to be error-free in the designated control slots.

First consider K=5 and 40 Hz or K=9 and 125 Hz based on the related/original (non-iterative) method, both give similar link performance (see Figs. 2 and 5), for a comparison between RS and CC coding methods. Fig. 10 shows the average probability of packet retransmission, as a function of occupancy for all available (24) channels in each sector. This is a measure of QoS (quality of service) experienced by individual users. With a 3-6 % target retransmission probability, 15-50 % occupancy per radio in each sector is possible with this DPA scheme, depending on the use of coding schemes. Clearly, joint channel estimation

and maximum likelihood detection of CC introduced previously provides significant improvement over the case of RS codes, which is also similar to the case of differential demodulation of the RS codes with 4 transmit antennas. Both results are significantly superior to the efficiency provided by current cellular systems, which are typically designed for voice communications with very conservative frequency reuse, about 4-7% spectrum occupancy in each sector. Data applications, permitting some retransmission delay, and improved link design, introduced here, allow much more aggressive and efficiency frequency reuse.

Fig. 11 shows that 1-1.5 Mb/s can be successfully delivered by each base station with an average delay on the order of 40-120 msec. This is a measure for system capacity. It indicates that OFDM link and DPA MAC combined enable a spectrally efficient (40%-60% b/s/Hz with a conservative assumption of overhead requirements) air interface for broadband services, even for the macrocellular environment considered here. Adaptive modulation has not been considered in this study, and its use is expected to improve efficiency beyond 1 b/s/Hz per base station even under aggressive frequency reuse. The OFDM technology discussed herein can provide robust performance with peak-rates scalable with the available bandwidth.

Next, consider the case of high maximum Doppler frequency (200 kHz) and K=9 (WER curves in Fig. 6) for comparison between sub-optimal ("original") detection method discussed earlier and near-optimal iterative backward-processing method presented. Fig. 12 shows that retransmission probability using the improved method can work well even under high maximum Doppler frequency. As a result, QoS can be improved even for high mobility users or when higher carrier frequency is employed. On the other hand, the capacity

difference is relatively smaller, as shown by the delay-throughput curves in Fig. 13. This is because both methods give very good radio link performance and the delay is dominated by sharing limited number of traffic slots, which is independent of the WER performance. In this case, better traffic resource management, such as improved admission control, could achieve capacity improvement. If higher Doppler frequency, e.g., 400 Hz, were encountered, the improved link performance introduced by the iterative backward-processing method, as shown in Fig. 8, would also result in system capacity enhancement.

Returning to Fig. 1A to show the detail of the decoder (DEC) unit and its relation with the Channel Estimator (CE) unit, with channel estimates $\hat{\mathbf{H}}_{m,n}$, $\hat{\mathbf{c}}_n$ can be calculated by maximum likelihood (ML) decoding (9),

$$\hat{\mathbf{c}}_n = \arg\min_{\mathbf{c}_n} \sum_m \|\mathbf{x}_{m,n} - \hat{\mathbf{H}}_{m,n}\mathbf{c}_n\|^2 .$$

In Fig. 1A, this decoding process is divided into 4 parts, e.g. QPSK Demodulator, a maximum ratio combiner (MRC), Deinterleaver and Viterbi decoder. How to separate these 4 units is now explained.

We have

$$\sum_m \|\mathbf{x}_{m,n} - \hat{\mathbf{H}}_{m,n}\mathbf{c}_n\|^2$$

$$= \sum_k (\sum_m \|x_{m,n,k} - h_{m,n,k}c_{n,k}\|^2)$$

$$= \sum_k (\sum_m \|x_{m,n,k}\|^2 - 2\operatorname{Re}\{x^*_{m,n,k}h_{m,n,k}c_{n,k}\} + \|h_{m,n,k}c_{n,k}\|^2)$$

Note that $\sum_{k}\sum_{m}\|x_{m,n,k}\|^2$ makes no contribution to our minimization and due to binary convolutional code with QPSK modulation, $\|c_{n,k}\|^2$ is a constant. $\sum_{k}(\sum_{m}\|h_{m,n,k}\|^2\|c_{n,k}\|^2)$ makes no contribution to our minimization either. So we concentrate on

$$\sum_{k}(\sum_{m}\mathrm{Re}\{x^*_{m,n,k}h_{m,n,k}c_{n,k}\}) = \sum_{k}\mathrm{Re}\{(\sum_{m}x^*_{m,n,k}h_{m,n,k})c_{n,k}\}$$

$$= \sum_{k}\left(\left[\sum_{m}\mathrm{Re}\{x^*_{m,n,k}h_{m,n,k}\}\right]\mathrm{Re}\{c_{n,k}\} - \left[\sum_{m}\mathrm{Im}\{x^*_{m,n,k}h_{m,n,k}\}\right]\mathrm{Im}\{c_{n,k}\}\right)$$

The demodulator and the MRC calculate $\sum_{m}\mathrm{Re}\{x^*_{m,n,k}h_{m,n,k}\}$ and $\sum_{m}\mathrm{Im}\{x^*_{m,n,k}h_{m,n,k}\}$.

And the remainder is done by the deinterleaver and the decoder.

Conceptually, the MRC does only energy combining. So, we perform the following calculations:

$$\sum_{m}\mathrm{Re}\{x^*_{m,n,k}h_{m,n,k}\} = \sum_{m}\mathrm{Re}\{x^*_{m,n,k}\arg(h_{m,n,k})\}\|h_{m,n,k}\|$$

$$\sum_{m}\mathrm{Im}\{x^*_{m,n,k}h_{m,n,k}\} = \sum_{m}\mathrm{Im}\{x^*_{m,n,k}\arg(h_{m,n,k})\}\|h_{m,n,k}\|.$$

Then, $\mathrm{Re}\{x^*_{m,n,k}\arg(h_{m,n,k})\}$ and $\mathrm{Im}\{x^*_{m,n,k}\arg(h_{m,n,k})\}$ are actually QPSK demodulation. And the remainder is the MRC.

The radio-link performance of our COFDM system with the novel and non-obvious combination of two channel estimation schemes under fast fading validate the design methodology and system for near optimal joint channel estimation and data detection.

For a COFDM system, coherent detection can significantly improve radio link performance, but it requires a channel estimator to achieve the potential gain. The present invention applies a forward channel estimation and maximum likelihood decoding method

for a COFDM system with convolutional codes and a simple interleaving scheme. It is shown that the performance of such a system significantly outperforms the one with Reed-Solomon codes. The performance under different fading rates was studied to understand the performance limits and areas for improvement. Simulations showed that this detection method is able to achieve near optimal performance in a wireless environment with maximum Doppler frequency as high as 100 Hz. However, as fading rate increases further, irreducible error floor is introduced by the limitation in channel-tracking capability. A near-optimal detection method has been presented to improve the channel tracking performance, which is based on iterative processing and iterative backward processing of channel estimation and data decoding. Within practical values of interest, error floor can be eliminated even for very high fading rates, which is important for cases with high mobility or when higher carrier frequencies are considered in the future. Radio system performance based on the improved link techniques and a medium access control protocol using dynamic packet assignment was also studied. With radio link improvement, system throughput and delay can be significantly enhanced for better quality of service and spectrum efficiency. The system is also greatly simplified by allowing reuse factor of one without frequency planning.

It should be clear from the foregoing that the objectives of the invention have been met. While particular embodiments of the present invention have been described and illustrated, it should be noted that the invention is not limited thereto since modifications may be made by persons skilled in the art. The present application contemplates any and all modifications that fall within the spirit and scope of the underlying invention disclosed and claimed herein.